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LOW INPUT VOLTAGE D. C. TO D. C. CONVERTER

REPORT NUMBER II

48 p.

Contract Number NAS 5-3441

National Aeronautics and Space Administration

Second Tertiary Progress Report

27 September 1963 to 26 December 1963

Goddard Space Flight Center

Greenbelt, Maryland

Submitted by

Minneapolis-Honeywell Regulator Company

ORDNANCE DIVISION

Hopkins, Minnesota

Honeywell

ORDNANCE

A DIVISION OF THE

MILITARY PRODUCTS GROUP

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Object: The object of this contract is to design and develop an efficient, reliable low voltage D. C. to high voltage D. C. regulated converter using germanium power transistor switching elements. This converter will be used to convert the low voltage levels of newly developed energy sources to useful higher levels.

Prepared by

John T. Lingle - Project Engineer

Submitted by

Minneapolis-Honeywell Regulator Company
ORDNANCE DIVISION
Hopkins, Minnesota

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SECTION I

PURPOSE

The purpose of this contract is to design and to develop an efficient, reliable, and lightweight, low voltage d. c. to high voltage d. c. regulated converter, using germanium transistors as the power switching element. The converter will be designed to convert the output of thermionic diodes, thermoelectric generators, fuel cells, solar cells, and high performance, single cell electrochemical batteries to a regulated 28 volt d. c. output. The program includes circuit configuration selection, optimization, and new design efforts to reduce losses, size and weight. Effort will be directed toward construction of a breadboard model to verify that the design goals and performance requirements have been optimized.

SECTION II

SUMMARY

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During the second quarter effort was directed toward simplification of the low input voltage converter regulator circuit, improving circuit performance, and providing overload protection. The synchronized power oscillator used previously has been replaced by a simpler circuit that has an operating frequency proportional to input voltage. This change was made for three reasons:

1. It was determined that it is not necessary to synchronize the regulator at some multiple of the converter frequency.

2. It is desirable to control the power transformer maximum flux density by making the converter operating frequency proportional to input voltage.

3. Simplification of the circuit reduces size and weight and increases reliability.

Breadboard tests with the simplified circuit have shown improved performance. Efficiencies of the regulated converter have been increased to between 74 and 76.5% for 50 watt outputs. There are indications that this efficiency can be increased by eliminating flexible leads between the transformer and the transistors and by selecting transistors with lower saturation voltages. Performance data to substantiate predictions of increased efficiency exists.

Preliminary ambient temperature tests have shown that the converter functions well over the required ambient temperature range, and also operates well at temperatures much lower than required.

Author

The use of low saturation voltage, high speed silicon transistors as pulse width modulation regulators is currently being investigated. The use of these transistors results in an all silicon regulator circuit. Preliminary results have shown a one to two per cent efficiency decrease with the all silicon chopping regulator; however, the increased temperature capability of silicon regulator circuitry may warrant its use in some applications.

The converter overload protection circuit has been incorporated, and checks show that it protects the device from slowly rising overloads as well as dead shorts. The converter recovers to normal operation when the overload is removed.

The work during this quarter has resulted in simplification of circuitry, improved performance, and incorporation of additional required circuit functions. Work must now be directed toward minimization of size and weight. Model design is an important factor in minimizing size and weight and in improving device efficiency by minimizing the length of high current conductors. Estimates based upon sound data place the weight and size of the present breadboard circuit (packaged in a container) at six pounds and 150 cubic inches. During the quarter, effort will be concentrated on model design and transformer design to reduce the size and weight. A cylindrical construction may be considered to minimize container weight.

SECTION III

CONFERENCES

Some phone conversations were made during this reporting period, but no formal conferences were held. It is anticipated that a conference will be scheduled in the next two weeks to discuss the details of this program, circuit parameters, and deliverable breadboard model fabrication.

SECTION IV

PROJECT DETAILS

A. CIRCUIT DESIGN

Circuit design effort has been directed towards simplification of the basic low input voltage converter, improvement of the voltage regulator, and incorporation of overload protection circuitry.

1. Low Input Voltage Converter-Regulator Block Diagram

The converter-regulator block diagram is shown on Figure 1. The converter section consists of a starting oscillator, to start the circuit under all conditions; a current feedback power oscillator, to chop the low voltage power and transform it to a higher voltage; a rectification circuit to change the square wave ac to dc; and a filter.

The voltage regulator consists of a transistor chopper, which efficiently chops and pulse width modulates the unregulated dc power. The pulse width modulated dc power then passes through a filter circuit which smooths the pulse power to the desired regulated dc voltage. The output voltage is sampled and compared with a zener diode reference by the differential detector circuit. The resulting amplified error signal is applied to a snap-acting trigger amplifier that switches the chopping power transistor, to pulse width modulate the dc power and close the regulator servo control loop. A saw tooth ramp generator supplies a dither signal to the differential detector circuit to control uniformly the frequency of the pulse width modulation circuit.

An overload protection circuit senses the output current and activates the snap acting amplifier and pulse width modulation transistors to limit the current to a small value during overload.

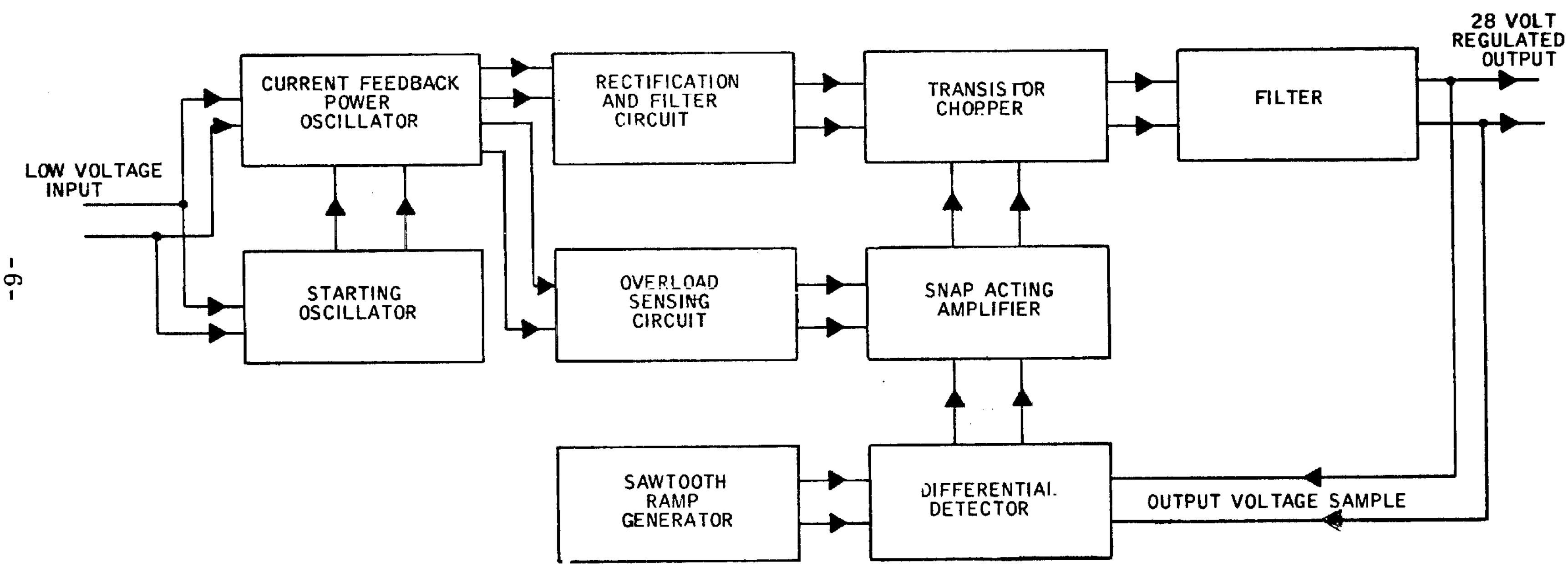


Figure 1 - BLOCK DIAGRAM

2. Circuit Configuration

The regulated 50 watt transistor low input voltage converter is shown on Figure 2. The circuit is similar to that shown in Progress Report Number 1, but there have been enough circuit changes to warrant a complete description of the new circuit.

Current Feedback Power Oscillator - The current feedback transistor power oscillator uses two Honeywell MHT 2205 low saturation voltage transistors in parallel in each push-pull half of the circuit. This current feedback power oscillator consists of transistors Q2, Q3, in one push-pull half, and transistors Q4, Q5 in the other push-pull half. The power oscillator also includes output power transformer T3, current feedback transformer T2, and the frequency controlling reactor L1. When the circuit is energized, it starts from either the dissimilarities in transistor leakage parameters, or by a pulse from the starting oscillator. During operation, one pair of transistors is rendered conductive and the other pair is back biased. Assuming that transistors Q4 and Q5 are rendered conductive, the current flow can be traced through the oscillator as follows:

Current flows through the positive input leads to the center tap of N1 on power transformer T3, then passes through winding N1B on transformer T3, to winding N2B on current feedback transformer T2. The current then passes through winding N2B, inducing magnetic flux in the core of transformer T2. This magnetomotive force induces voltages in T2, windings N1A, N1B, N2B, N2A and N3. The current flowing through winding N2B on T2 then passes to the emitters of transistors Q4 and Q5.

Since these are rendered conductive, the current flows through the Q4 and Q5 emitter-collector junctions to the negative input line, completing the primary

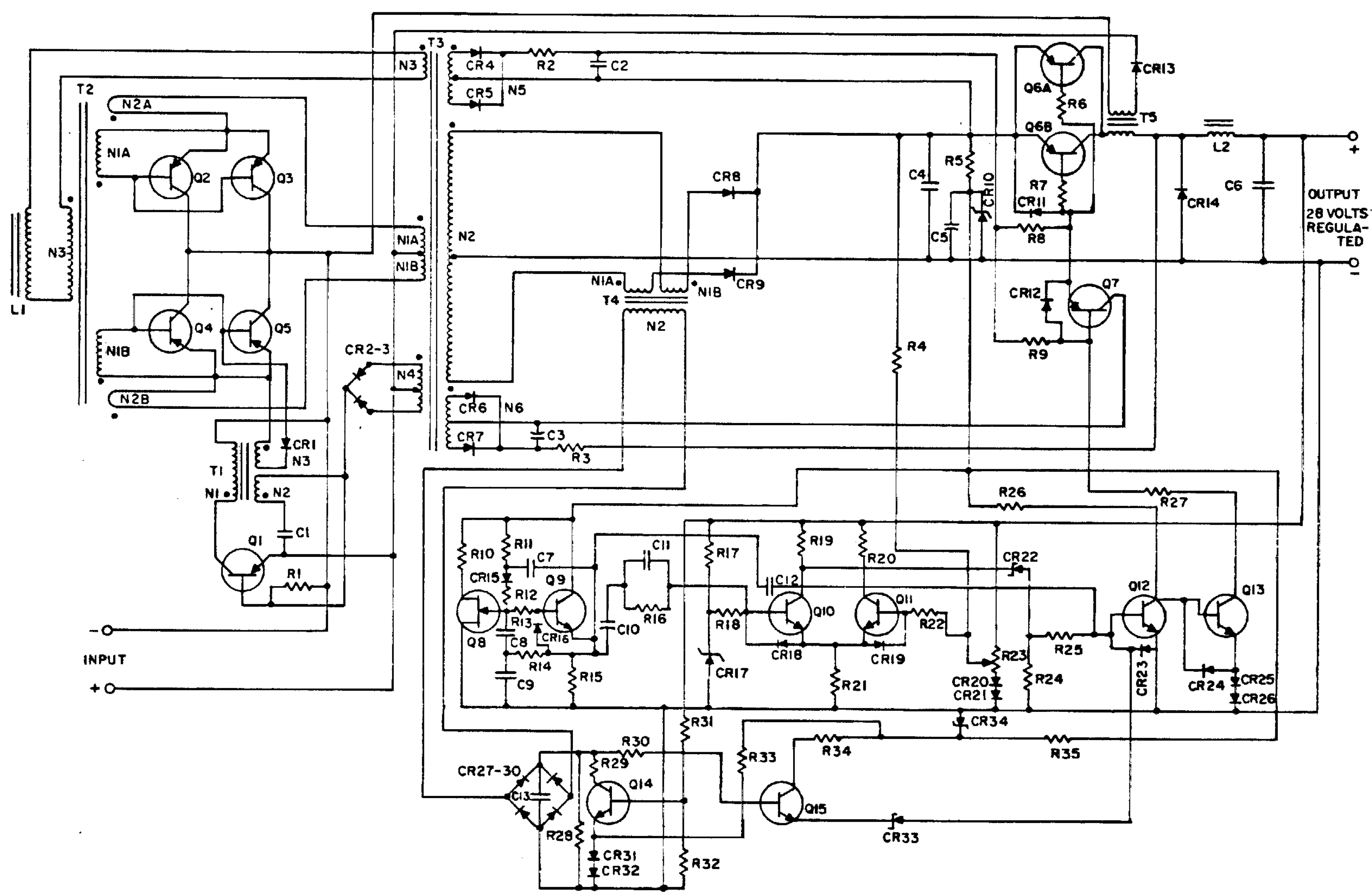


Figure 2 - CIRCUIT DIAGRAM

circuit for one push-pull half of the power transistor oscillator. The magnetic flux in current feedback transformer T2 induces the polarities indicated by the dots in Figure 2. At this instant T2 flux causes the lower portion of N1B to be positive and the upper portion to be negative. The lower portion is connected to the emitters of Q4, Q5, and the upper portion is connected to the bases of Q4 and Q5. This biases the base negative, renders transistors Q4, Q5 conductive, and drives them into saturation so they can pass the heavy primary current from emitter to collector. Thus the current flow through winding N2B on T2, induces voltage and current in winding N1B, which provides positive feedback for the oscillator, and drives transistors Q4, Q5 into conduction. At the same time, the voltage induced in winding N1A on T2 applies a positive potential to the bases of Q2, Q3 back biasing them "off". Thus during this half cycle, transistors Q4, Q5 are rendered conductive and transistors Q2, Q3 are back biased by positive feedback through transformer T2.

The primary current flow through transformer T3 winding N1B induces the polarity shown by the dot on Figure 2. This induces voltages in T3 windings N1, N2, N3, N4, N5, and N6 to power the output circuit and operate the regulator small signal circuitry. A voltage is induced in winding N3 of power transformer T3, which is applied through a reactor L1 to winding N3 on current feedback transformer T2. The return lead goes to the other end of winding N3 on transformer T3. This current path applies negative feedback to the power oscillator. The inductive reactance of reactor L1, however, blocks nearly all of this negative feed-back for the major portion of the half cycle. Reactor L1 inductance limits the amount of negative feedback to a negligible level as long as L1 is unsaturated. However, after a time, inductor L1 will saturate and its impedance falls to nearly zero. When this occurs, a large amount of negative feedback is applied to winding N3 on current feedback transformer T2. This large negative feedback overpowers the positive feedback signal in T2, causing the conducting transistors to become non-conductive. This action recycles the power oscillator. The purpose of this negative feedback path is to control the operating frequency of the current feedback power oscillator. This arrangement causes the operating frequency to be proportional to input voltage. If the negative feedback route through T3 winding N3, reactor L1, and winding N3 on T2 were not used,then the frequency of the current feedback power oscillator would depend upon the load instead of input voltage. Frequency control proportional to voltage has been used so that the operating flux density in the power transformer T3 can be controlled to maximize efficiency and to prevent saturation of the output transformer at high voltages and light loads. This also tends to minimize saturation of the output transformer due to half cycle unbalance.

The current feedback transistor oscillator is generally self starting; however, to insure starting under all environmental conditions a separate starting oscillator has been incorporated to provide pulses that will guarantee starting under all environmental and load conditions. The starting circuit is a blocking oscillator consisting of transistor Q1, transformer T1, diode CR1, capacitor C1, and resistor R1. The emitter of transistor Q1 is connected to the positive line. The base of transistor Q1 is connected to the negative line through resistor R1 and to the positive line through a turn off network consisting of winding N4 on transformer T3 and diodes CR2 and CR3. Capacitor C1 and winding N2 on T1 are also connected across the emitter base circuit. The collector is connected to winding N1 on T1, and the other end of winding N1 is connected to the negative input line.

When power is applied to the input line, current flows to the emitter of transistor Q1 and through its base circuit through resistor R1 to the negative line. This renders transistor Q1 conductive, and current passes from emitter to collector through winding N1, energizing T1 and inducing voltages in windings

N1, N2, and N3. This momentarily applies a positive voltage to the T1 windings as indicated by the dots shown in Figure 2.

Induced voltage in winding N2 now applies a negative potential to the base of Q1 with respect to the emitter of Q1 through capacitor C1. The induced voltage in winding N2 momentarily provides positive feedback to transistor Q1, through capacitor C1, winding N2, and the base of Q1. This drives Q1 into the saturation region and allows a considerable amount of current to pass through winding N1. The resulting current flow induces sufficient voltage in winding N3 to draw current from the bases of Q4, Q5 through diode CR1 to T1 winding N3 which returns to the emitters of Q4, Q5. This renders transistors Q4, Q5 conductive and assures starting of the main current feedback power oscillator circuit. The Q1 blocking oscillator will continue to put out a series of pulses until they start the main power oscillator. Oscillation of Q1 is insured by the feedback path through winding N2 and C1. When C1 charges up, it can no longer pass the required drive current, and Q1 blocks, causing the transformer to reset. This recycles the blocking oscillator causing it to put out a series of starting pulses. After the main power oscillator starts, the voltage induced in power transformer T3 winding N4 applies a positive potential to the base of Q1 turning it "off." This renders the starting circuit non-operative as long as the main power oscillator functions. If for any reason the main power oscillator should stop operating the starting circuit would again oscillate and restart the main power oscillator. The starting oscillator becomes operational when a few tenths of a volt are applied to the input terminal.

Secondary Circuit - The voltage induced in winding N2 of power transformer T3, provides the stepped up power to operate the voltage regulator and drive the load. Secondary winding N2 is center tapped, and the output current flows through either half of winding N2 through a winding (N1A or N1B) on overload current sensing transformer T4 to one of the power rectifiers CR8 or CR9. This full wave rectifier then applies a positive 28 volts or more to filter capacitor C4 and the emitters of the chopping regulator transistors Q6A and Q6B. The negative output lead is connected to the center tap of winding N2 on T3. Thus the power amplifier chops the low voltage, high current dc power to square wave ac, and then raises it to a higher voltage. The rectification circuit rectifies this higher voltage ac power to dc and it is filtered by capacitor C4 and applied to the input of the voltage regulator circuit.

3. Voltage Regulator - The basic function of the voltage regulator was explained in the Block Diagram discussion above. Specific voltage regulator details are explained below:

The right side of Figure 2 shows the voltage regulator circuit. The converter rectified output is smoothed by filter capacitor C4, providing smooth unregulated dc power to the pulse width modulation regulator. The main function of this filter is to isolate the converter from the pulse width modulator. This isolation prevents pulse width modulator noise from being fed back onto the input line. Also, the converter power transistors conduct a uniform amount of current during each half cycle for the entire half and this maximizes the converter efficiency.

Transistors Q6A and Q6B chop the dc current to effectively pulse width modulate the flow of power to the output filter. The output filter consisting of L2, CR14, and C6 smooths the pulse power to regulated dc. To minimize power dissipation in the chopping transistor during switching, the reactance of a small transformer T5 has been placed in series with Q6A and Q6B. The operating characteristics of this circuit are such that when switching "on" the transistor must conduct all of the L2 choke current before it can take over from the freewheeling diode, CR14 and begin raising the choke terminal voltage. The transistor must conduct the entire load current before its emitter-to-collector voltage can drop to zero. This characteristic causes high instantaneous transistor switching dissipation.

The inductive winding of transformer T5 is placed in the transistor collector circuit to absorb an appreciable percentage of the instantaneous voltage drop when the transistor is initially switched into conduction. During this interval, the inductive voltage drop $e = -L \frac{di}{dt}$ will be large because of the rapid current increase. This reduces the transistor voltage drop considerably during this interval, and a quantity of energy is stored in the transformer's magnetic field in place of being dissipated in the transistor. The stored energy is fed back onto the input line during the resetting period through diode CR13. The inductance of this transformer is small, and hence it has little effect upon the quiescent portion of the conducting half cycle because the current rate of change is small. The effects of this element were previously described in Progress Report Number 1 in more detail.

Differential Detector and Pulse Width Modulator

The regulator output voltage is sampled by a potentiometer R23 across the output. This sample voltage is applied through R22 to one side of the differential detector at the base of Q11. The other side of the differential detector is connected to a zener diode reference CR17 through resistor R18. If the output voltage rises so that the voltage sample applied to the base of Q11 exceeds the reference voltage at the base of Q10, then a positive amplified error signal will appear at the collector of Q10. This positive error signal is applied through zener CR22 and R25 to the base of amplifier transistor Q12 and will tend to switch it into conduction. Q12 is normally held off by resistor R24, and requires a positive signal above a threshold value to cause it to conduct. When Q12 conducts, the voltage at the base of Q13 declines to 1.0 volt due to current flow through R26. Since the emitter of Q13 is biased to plus 1.4 volts by diodes CR25 and CR26, transistor Q12 will be back biased "off," removing the forward base drive from Q7, which is then back biased "off." A positive bias voltage higher than the converter output is provided by a small bias winding (T3, N5), rectifiers (CR 4, CR5) and filter (R2, C2). This bias winding applies a positive voltage source to back bias Q6A, Q6B, and Q7 through resistors R8 and R9, respectively. The bases of the regulator chopping transistors Q6A, Q6B are connected through R6 and R7 to the emitter of Q7, and thus Q6A and Q6B are back biased "off" when Q7 is back biased. Note that the differential detector output controls the chopping transistor through a network of three amplifying transistors. The resulting high gain insures rapid switching under all load conditions.

Regulator Operation

When the converter begins operation from a below normal voltage source, or from a slowly rising voltage source, the converter output voltage rises and a positive voltage is applied through resistors R5 and R26 to the base of Q13 turning it "on". This voltage causes transistors Q7, Q6A, and Q6B to conduct heavily so that power can pass through to the output circuit. With a low output voltage the error signal will be too low to initiate the chopping regulator, and it will remain fully conducting. This conduction will insure maximum output voltage for a low source voltage and heavy load conditions. As the output voltage rises toward the required value, the output voltage sample will approach that of the zener diode reference CR17. When this feedback voltage sample exceeds the reference, the amplified error signal will switch the three stage amplifier and will rapidly shut "off" the chopping transistors Q6A, Q6B. This action will stop power flow to the output filter and the output voltage will begin dropping. A slight output voltage drop will be sensed by the differential detector, and it will reverse the error signal to rapidly switch the chopping transistor back into conduction. Thus, the differential detector will cause the regulator to pulse width modulate, and the operating frequency and period will depend upon the load and the filter time constant. To get a more uniform regulator operating frequency a dither oscillator has been used.

The dither oscillator consists of a linear unijunction sawtooth generator. The unijunction relaxation oscillator fires when capacitors C8 and C9 become charged to a voltage sufficient to fire Q8. The relaxation oscillator output is applied to the base of an emitter follower consisting of Q9 and R15. A boot strap arrangement consisting of R11, R12, CR15, C7, and C8 is incorporated in this sawtooth generator to give a very linear sawtooth ramp for uniform operation of the switching regulator. More detailed information on this relaxation oscillator can be found in the General Electric Transistor Manual, Sixth Edition, p. 96. The linear sawtooth output of the dither oscillator is applied to the base of differential detector transistor, Q10, through a coupling circuit consisting of C10, C11, and R16. This R-C coupling circuit shapes the sawtooth wave to produce a very sharp spike on the leading edge. This sharp spike tends to operate the chopping regulator at a uniform frequency as the output voltage rises to the point where the regulator begins to initiate pulse width modulation. The linear ramp behind the spike tends to space uniformly the duration of the chopping transistor "off" time. The length of this "off" time will depend upon the magnitude of the ramp signal and the magnitude of the error signal. The differential detector will switch when the sum of reference and ramp signals equal the output voltage sample. Circuit operation and pulse duration time depend upon a coincidence in the ramp signal and output voltage sample magnitudes.

Overload Protection Circuit - Current transformer T4 has its primary winding split into two sections (N1A, N1B), and these are each connected in series with one half of the power transformer push-pull output. The primary winding has a few turns, and the secondary winding has many turns to provide a very high transformation ratio. The secondary of this current transformer is applied to the bridge rectifier, consisting of diodes CR27, 28, 29, and 30. Thus, current transformer T4 samples the amount of current flowing in the power secondary circuit and provides a small output current proportional to the load current flowing. The rectified current sample is filtered by capacitor C13 and is applied across transistor Q14. An output voltage sample is applied to the base of transistor Q14 through a network of resistors R31 and R32. This network will supply sufficient Q14 base current to render it conductive for a relatively small fixed amount of collector current. The current sample taken from T4 is small, below full load; consequently, transistor Q14 can conduct this without difficulty. The base current provided through resistor R31 is sufficient to maintain conduction for normal secondary load currents. However, when an overload or short circuit is applied to the converter output, the current flowing through transformer T4 increases and this increases the current through the secondary winding of T4, and transistor Q14 attempts to conduct more current.

When the output is overloaded or shorted, the output voltage tends to decline and the amount of current drive for transistor Q14 decreases due to the lower voltage across resistor R31. Thus during overload, transistor Q14 must conduct more current, yet its base drive through R31 tends to decline. This increases the saturation voltage drop across Q14 under overload conditions. As the overload increases, the ability of the transistor to conduct heavier currents diminishes, causing the voltage drop across Q14 to increase. Overload increases the potential of resistor R30 and causes current to flow into the base of transistor Q15 rendering it conductive. A path from the power oscillator rectified output through resistor R5, resistor R35, and resistor R34 applies potential to the collector of transistor Q15. The emitter of transistor Q15 is connected through Zener diode CR33 to the base of transistor Q12. When the base of transistor Q15 becomes sufficiently positive to override the base emitter voltage drop of Q15 and the zener voltage of CR33, current will flow through Q15 and CR33 to the base of Q12 and render Q12 conductive. Thus an overload will cause transistor Q12 to conduct. This in turn back biases transistors Q13, Q7, and the chopping regulator transistors Q6A and Q6B. Thus an overload will shut "off" the regulator chopping transistors to limit the output current to a safe value. This prevents the converter and the converter regulator from damage by overloads or short circuits.

If the overload is removed, the signal to maintain the regulator non-conductive will also be removed because the output voltage will tend to rise and current will again flow through resistor R31 to forward bias Q14. Conduction of Q14 will again remove the drive signal from transistor Q15 and allow the regulator to again conduct heavily. Thus a short circuit will cause the overload circuit to trip, and this will then shut off the chopping regulator transistors limiting the load current or short circuit current to a safe value. When the short or overload is removed, the regulator circuit will again conduct normally and provide useful output to the loads. The characteristics of this overload protection circuit are such that it will protect for slowly rising overloads, for moderate overloads of a fixed value, or for dead shorts.

The circuit will limit overload current to a fixed value and prevent damage to any of the converter components. A more complete description of the operating characteristics of this circuit are contained in a paper on overload protection presented at the 16th Annual Power Sources Conference by J. L. Jensen. For this circuit to provide a small amount of output current during an overload or a short circuit condition, a path is provided from the dither oscillator through capacitor C12 to apply a small dither signal to the base of Q12. This dither signal continues to pulse the chopping regulator transistors Q6A and Q6B into conduction for a very small portion of the cycle so that a small limited output current flows to the load. This short circuit current is used to sense the state of the output load. Thus, when the overload or short is removed, it causes the output voltage to rise to the point where it can deactivate the overload protection circuit and again allow full power to be supplied to the normal type of load impedance.

B. CONVERTER PERFORMANCE

1. Preliminary Ambient Temperature Tests

The converter breadboard was subjected to preliminary ambient temperature tests to determine which areas would need further modification before the design became more finalized. The converter regulator was checked at ambients of 70°C, 25°C, 0°C, -10°C, and -54°C. During these tests the blocking oscillator circuit was modified to insure that it would operate and be capable of starting the current feedback power oscillator under all conditions. The starting oscillator circuit, shown on Figure 2, was found to work best and guarantee starting of the power oscillator over the above temperature range. The device was checked at lower temperatures than required (-54°C compared with -10°C required) because we thought that it would function at these temperatures and testing to lower temperature limits would guarantee removal of any marginal modes of operation. Thus, a circuit tested to lower limits would tend to indicate greater reliability because it would be less susceptible to variations and changes in component parameters.

At very low temperatures (-54°C) the pulse width modulation transistors Q6A and Q6B were found to come out of saturation toward the end of the conducting period and cause a minor oscillation. This difficulty was traced to the dither oscillator which was functioning improperly. Diode CR15 was added to the circuit, and the parameters were adjusted to remedy this condition. This change produced the proper wave shapes for operation of the regulator under the above ambient temperature conditions.

Ambient temperature tests were conducted to check the output voltage regulation over the temperature range. Diodes CR20 and CR21 were added to the error sensing circuit to compensate the regulator for temperature changes. Curves showing output voltage vs load for a 1.2 volt input and various temperatures are shown on Figure 3. There was a slight tendency for the output voltage to rise at the -10°C temperature. The voltage regulation is within the ± 1% requirement over the temperature range.

Performance curves of regulated converter efficiency vs power output were also plotted for ambient temperatures of 70°C, 25°C, -10°C, and -54°C. These are shown in Figures 4, 5, 6 and 7, respectively. The maximum efficiency for each temperature exceeds 70%. It can be noted that in general the overall efficiency increases slightly with a drop in temperature. This can be attributed to a decrease in transformer winding resistance and to characteristics of the Honeywell MHT 2205 transistors which exhibit an increase in gain at lower temperatures.

After the preliminary temperature tests and modifications showed satisfactory results, further effort was directed towards improving device performance. A new power transformer was designed and parameters of the regulator circuit parameters were changed. The overload protection circuit was also incorporated into the regulator. Room temperature performance data was then recorded on the complete converter-regulator.

2. Modified Converter-Regulator Performance

The complete and modified converter-regulator showed improved performance. The overall efficiency vs load curves for various input voltages are shown on Figure 8. The overall efficiency exceeded 70% for inputs between the 1.2 volt and the 1.6 volt range. At 1.2 volts the efficiency was above 70% for power outputs between 19 and 55 watts. For this same input voltage the efficiency exceeded 75% over the 34 to 55 watt range. The maximum efficiency was 76.5%

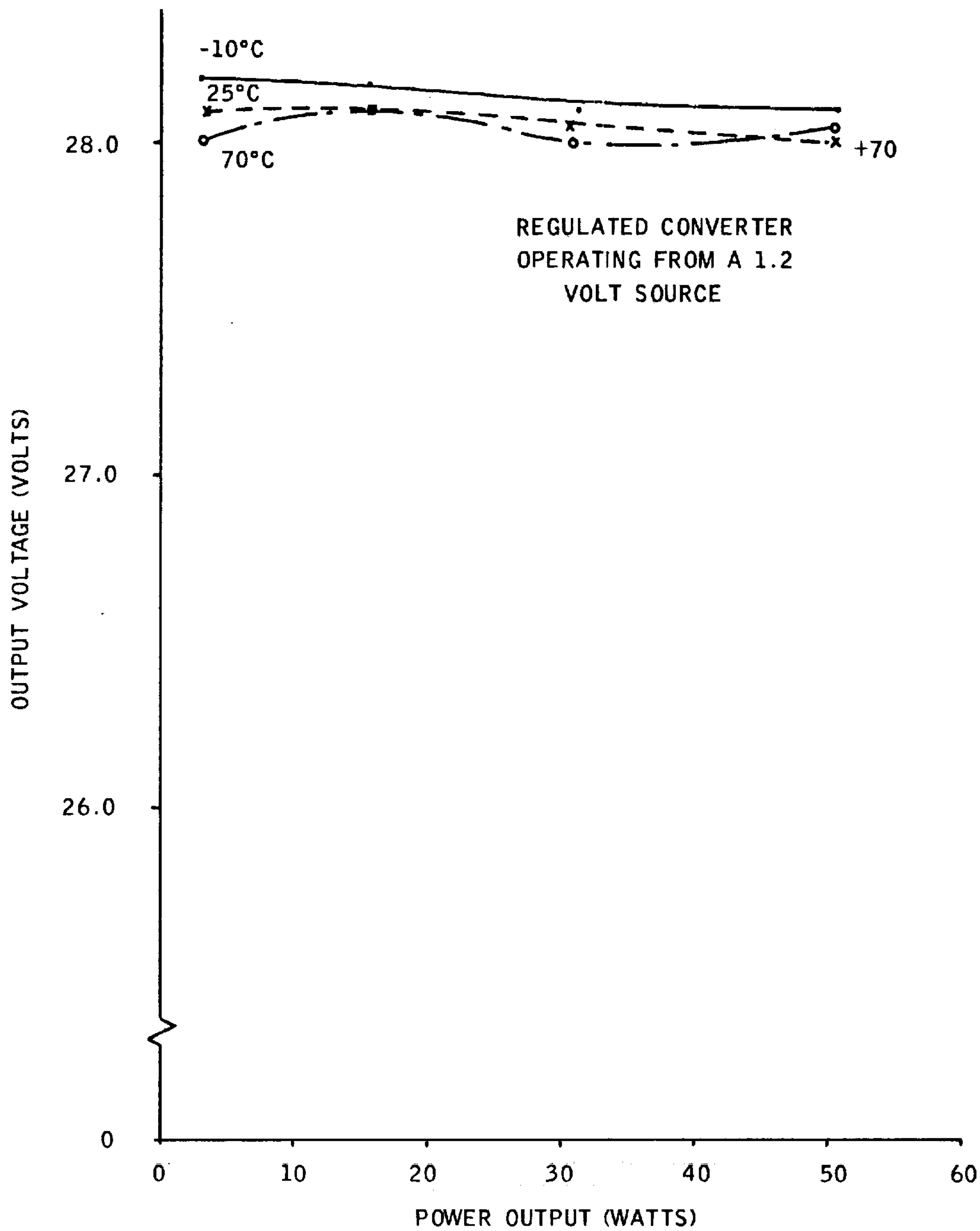


Figure 3 - OUTPUT VOLTAGE VS LOAD AT VARIOUS TEMPERATURES

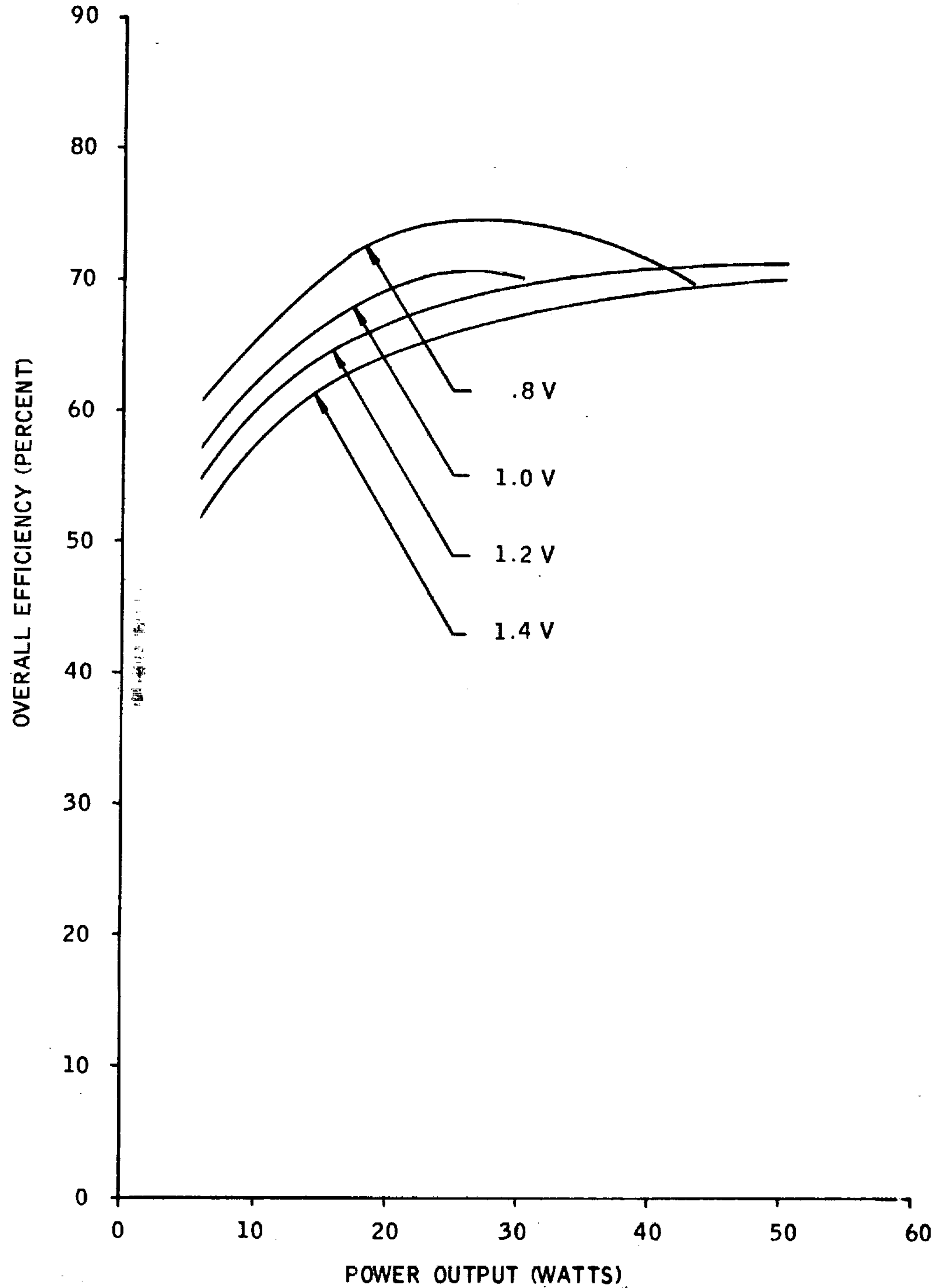


Figure 4 - REGULATED CONVERTER EFFICIENCY VS POWER OUTPUT AT 70°C AMBIENT

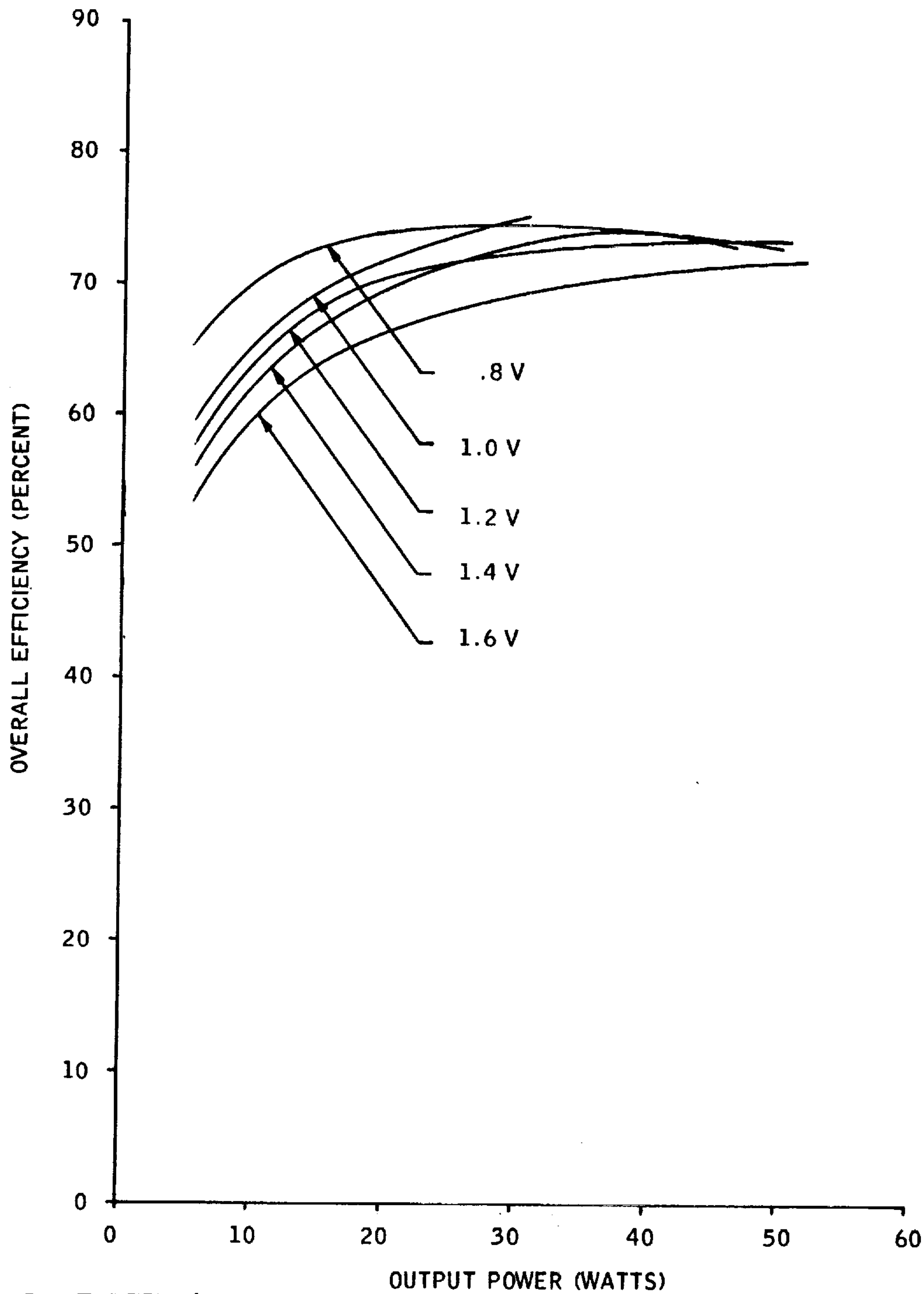


Figure 5 - REGULATED CONVERTER EFFICIENCY VS POWER OUTPUT AT 25° C AMBIENT

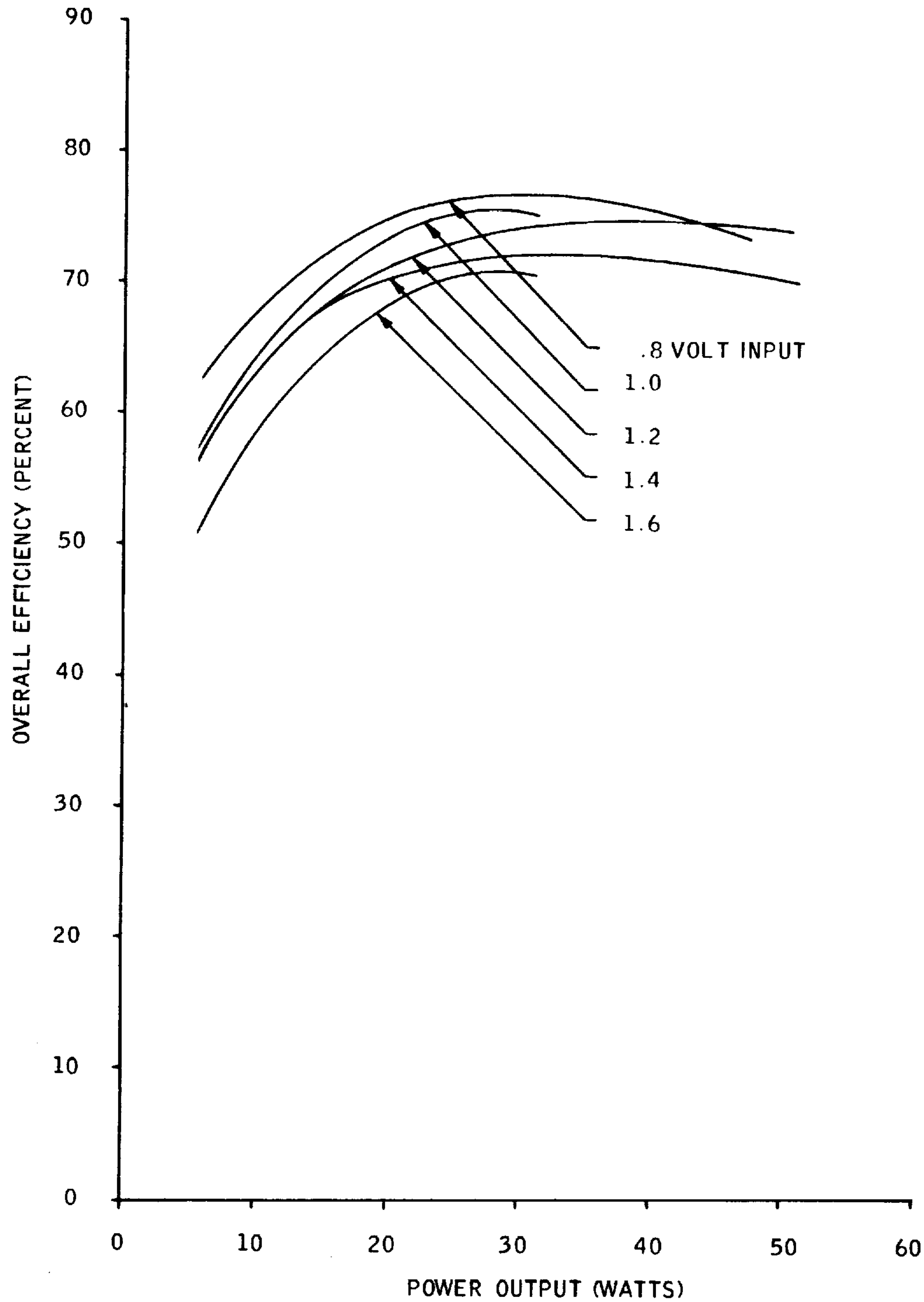


Figure 6 - EFFICIENCY VS POWER OUTPUT AT -10°C

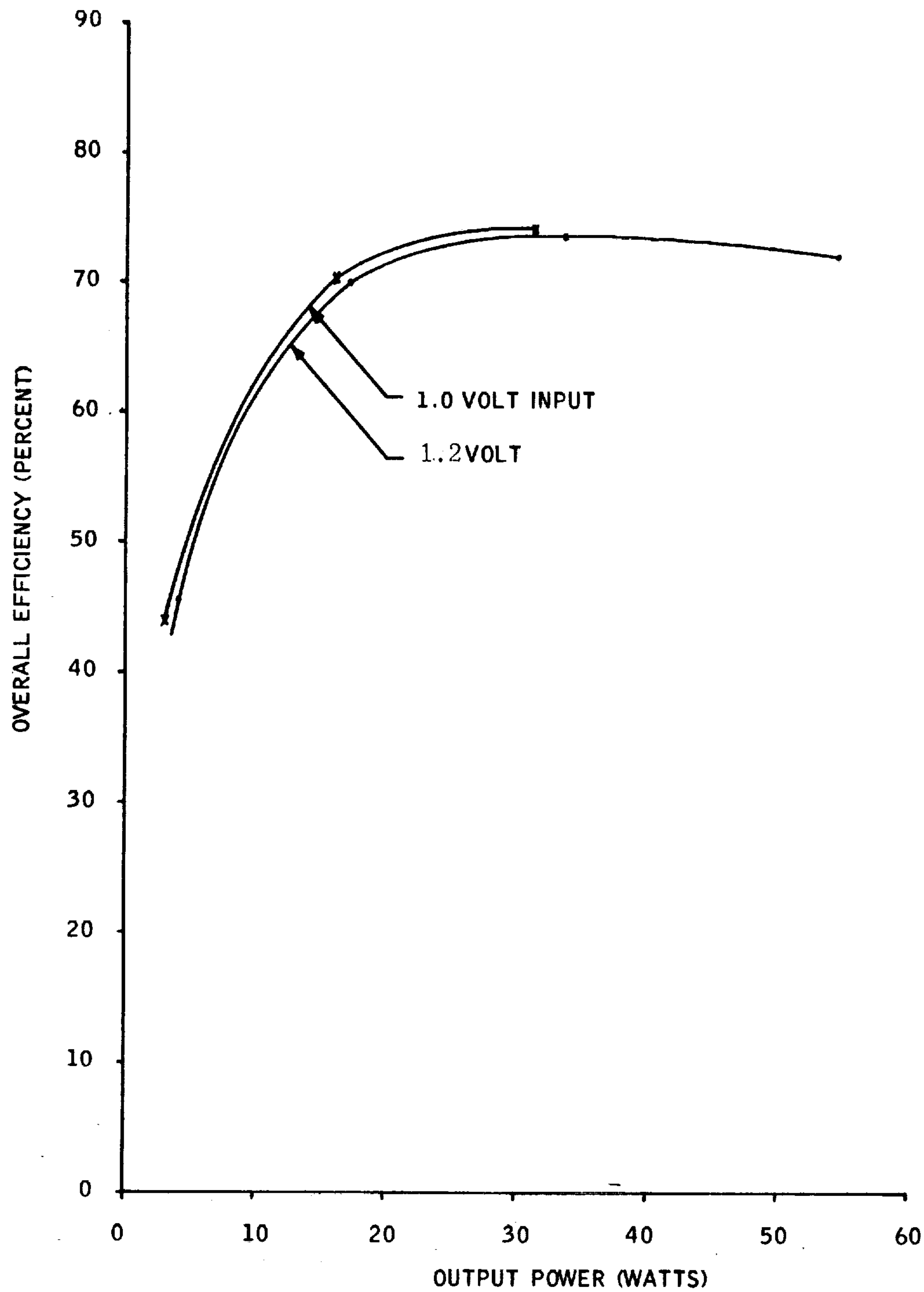


Figure 7 - EFFICIENCY VS POWER OUTPUT AT -54°C

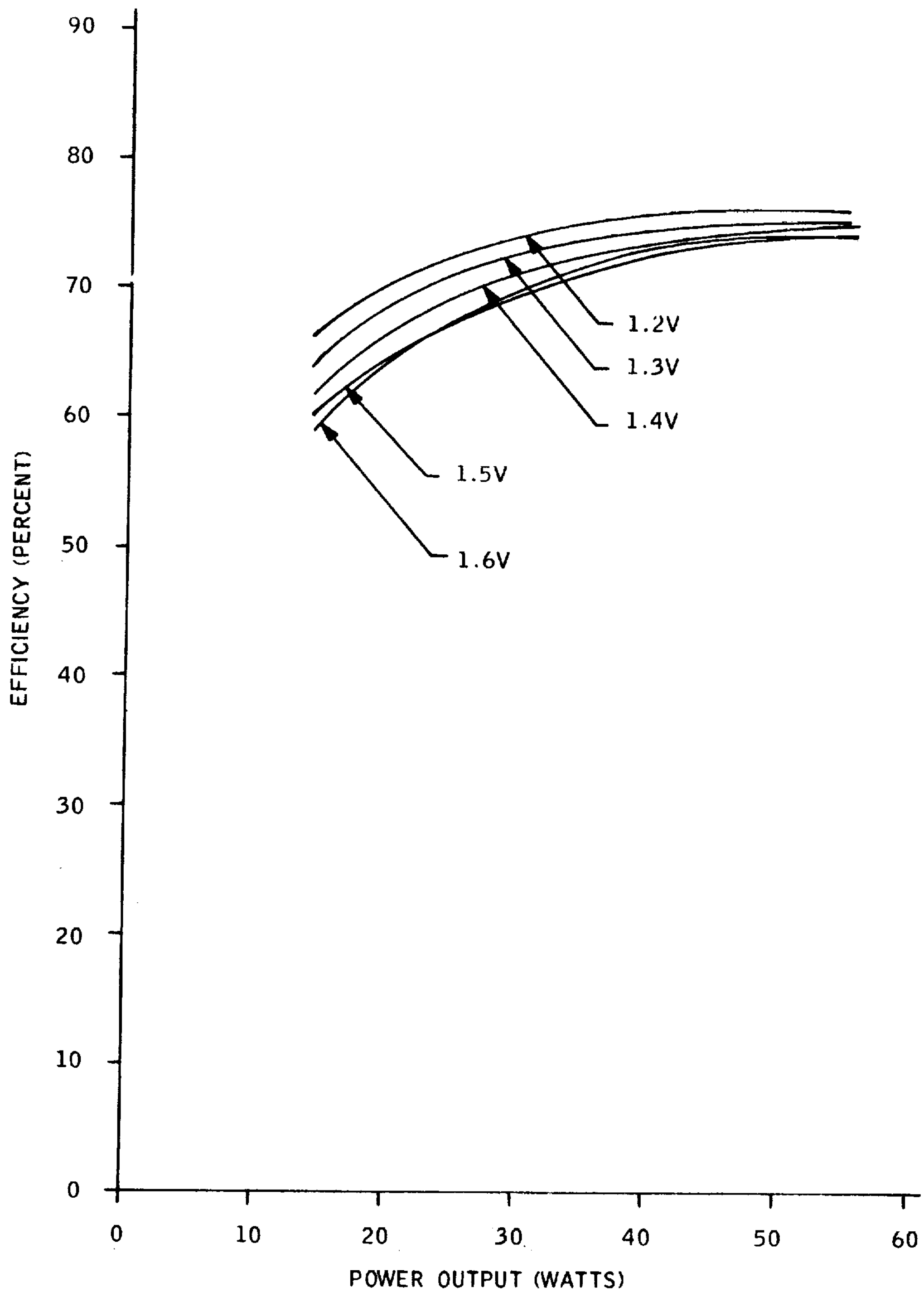


Figure 8 - OVERALL EFFICIENCY VS OUTPUT POWER FOR THE IMPROVED CONVERTER-REGULATOR

at 55 watts load. The curves for higher input voltages tended to follow the same pattern. The maximum efficiency for a 1.6 volt input was 74.5% at 55 watts output. The efficiency remained above 70% over the 36 to 55 watt range for a 1.6 volt input. The family of performance curves was uniform and was spaced between the 1.2 volt and 1.6 volt curves.

The efficiencies were lower for the higher input voltages. This efficiency decline is more predominant at light loads. The decline in efficiency with higher input voltages is caused by a lower conduction period in the pulse width modulation regulator. Thus for a given power output, the power must flow through the regulator in a shorter time if the input voltage is high. This increases the choke core loss (L2), regulator transistor switching loss (Q6A, Q6B), and the free wheeling diode dissipation (CR14).

The reduction in regulator conduction time for higher input voltages is shown on Figure 9, which shows the regulator "Percent of Conduction Time vs Output Power" for various input voltages. For the 1.2 volt input, the per cent of conduction time increased from 91% to 97% over the 15 to 55 watt range. For higher input voltages, the percent of conduction time decreased. For the 1.6 volt input the conduction time ranged between 67% and 70% over the 15 to 55 watt range. It can be seen that increasing the voltage input swing will decrease regulator efficiency at high input voltages.

The overall converter-regulator efficiency is a product of the converter efficiency and the regulator efficiency. It has been shown that the regulator efficiency declines with higher input voltages. In general, the converse is true with the converter since its efficiency usually tends to increase with higher input voltages at heavy load. This is not always true at light load because transformer core losses are higher with higher input voltages. Thus input voltage effects tend to cancel each other in the overall converter-regulator

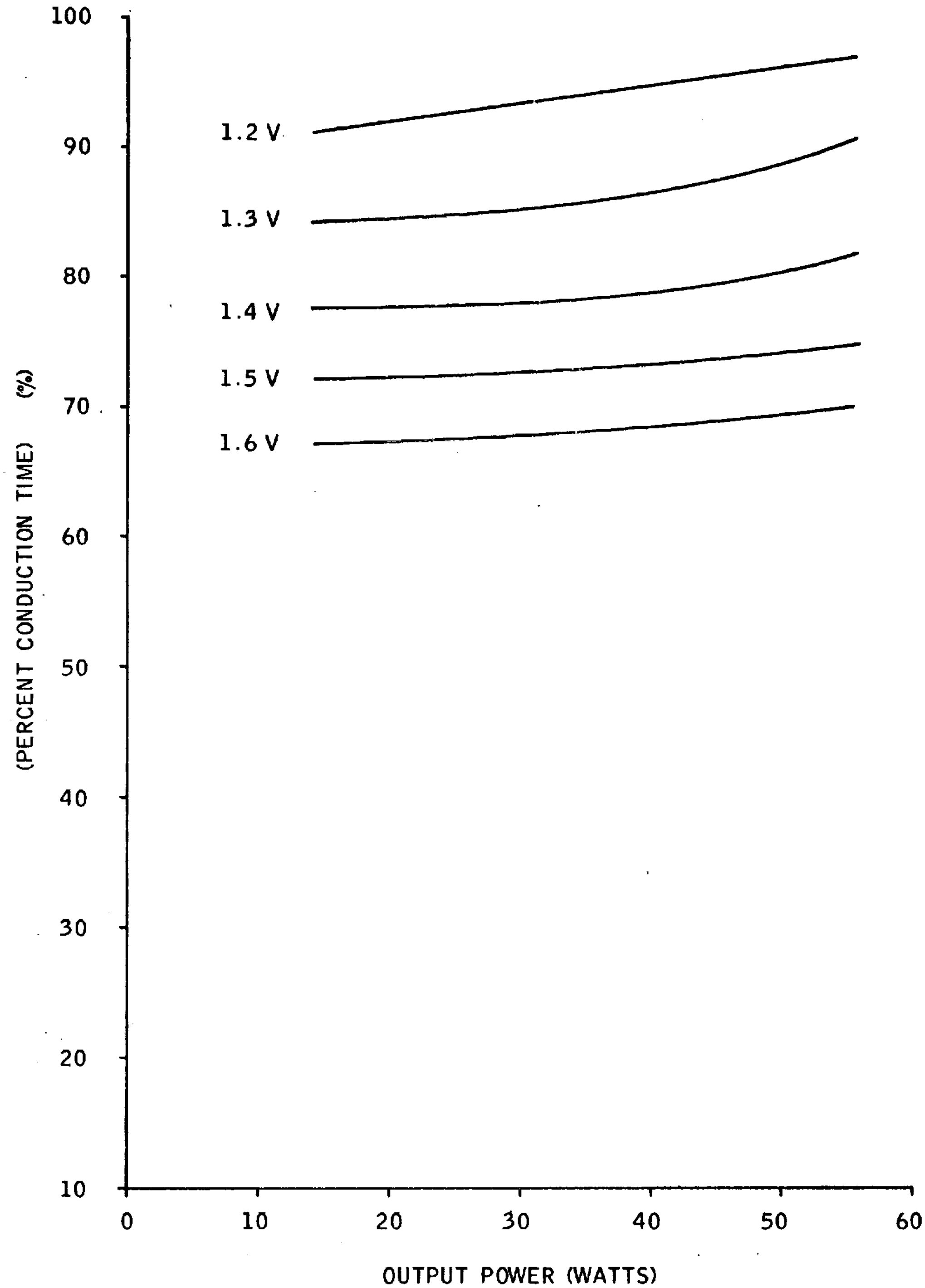


Figure 9 - REGULATOR PER CENT CONDUCTION TIME VS OUTPUT POWER

system, which accounts for the narrow efficiency variation over the 1.2 to 1.6 volt input range at 55 watts load on Figure 8. The efficiency spread is much larger at light load because both the converter and regulator efficiency tend to decline with higher input voltages at light load. With high input voltages the converter efficiency is less at light load because the transformer core losses are slightly larger.

The voltage regulation characteristics of the converter-regulator are shown on Figure 10. Note that for input voltages above 1.2 volts the output voltage remains quite flat over the load range. The output tends to droop with load at the 1.2 volt input because the regulator conduction time is approaching the maximum which is slightly less than 100 percent. The regulator gain declines near this limit, and the system IR drops cause the output voltage to decline with load. The family of curves shows that the regulator needs additional compensation for input voltage because the output rises slightly with higher input voltages. Despite these variations, the curves show that the regulator holds the output voltage to within the ± 1% required over the 15 to 50 watt load range shown.

The converter-regulator performance data shown on Figures 8, 9 and 10 described above utilized a pulse width modulation voltage regulator that used two high frequency germanium 2N2833 transistors in parallel for the regulator power switching element. It was necessary to use two 2N2833 transistors in parallel (Q6A and Q6B in Figure 2) to provide adequate power capability for transients during overload and short circuit tests.

3. Overload Protection

The overload protection circuit has been checked for operation with slowly rising overloads and for sudden short circuits. The circuit triggers on overloads in excess of 180% load. When the overload is removed, the overload

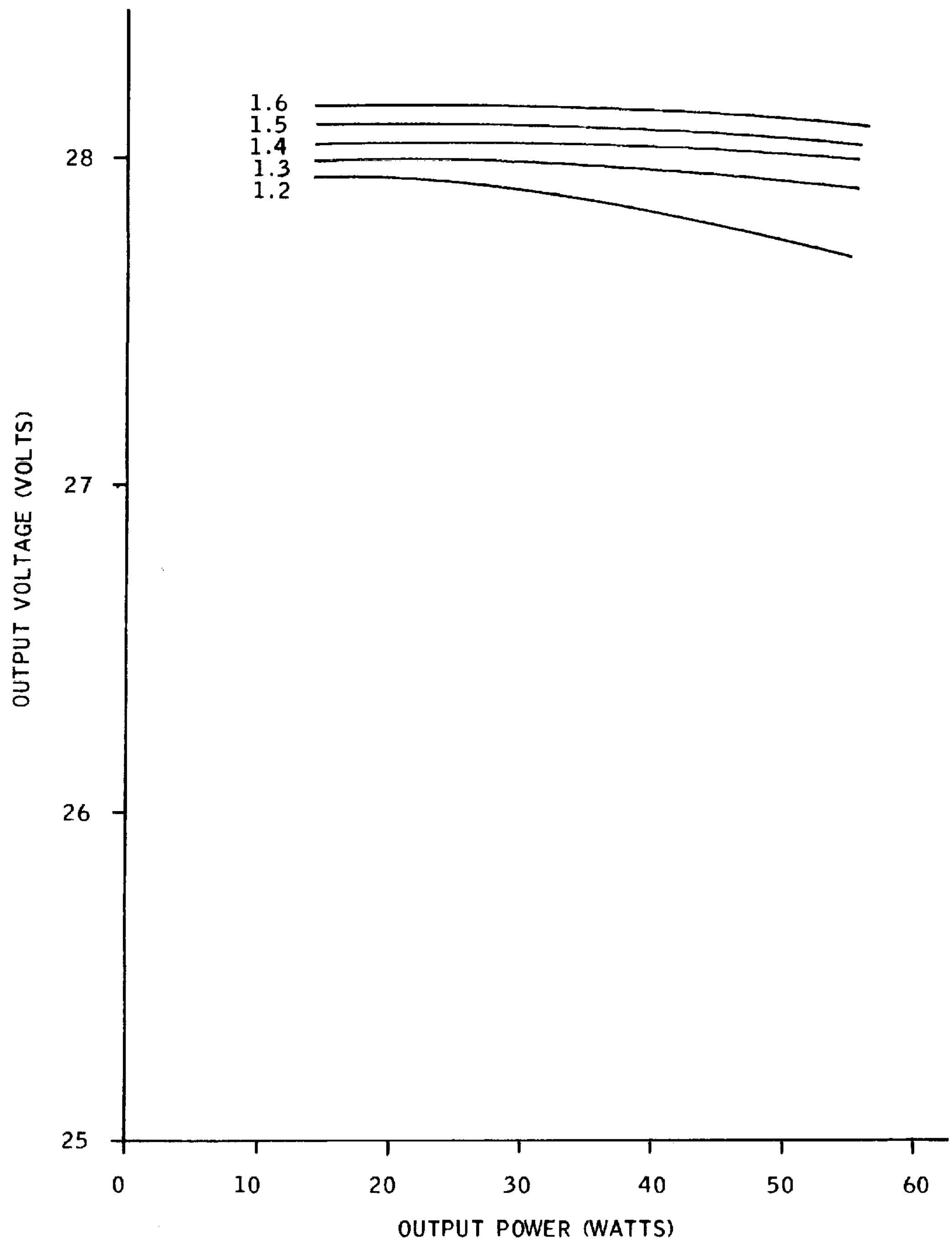


Figure 10 - OUTPUT VOLTAGE VS LOAD

protection circuit deactivates automatically, and the converter supplies full power to the normal load. Further tests must be conducted on this circuit to verify that operation is satisfactory for all environmental conditions.

4. Converter-Regulator Performance with an "All Silicon" Voltage Regulator

Since the germanium transistors have lower temperature limits than silicon transistors, the performance of the regulator utilizing all silicon transistors was determined. A new high frequency 20 ampere, planar triple-diffused, low saturation voltage, Honeywell MHT 8003 silicon transistor was incorporated into the pulse width modulation regulator. The silicon transistor has a slightly higher saturation voltage than the germanium 2N2833, but these newer planar triple-diffused silicon devices have much lower saturation voltages than the 0.7 volt V_{CE} of previous silicon transistors. An MHT 4518 silicon transistor was substituted for Q7 to drive the MHT 8003 used in place of Q6A and Q6B in Figure 2. The performance curves for the low input voltage converter with an "all silicon" voltage regulator are shown on Figures 11, 12 and 13.

The "Overall Efficiency vs Output Power" curves on Figure 11 show that high overall efficiencies were obtained with an all silicon voltage regulator. The curves show that for a 1.2 volt input the efficiency remained over 70% for loads between 19 and 55 watts. For the 1.6 volt input the efficiency was over 70% for loads between 32.5 and 55 watts. The curves have the same general shape as those for the germanium regulator shown in Figure 8. The only difference is the slight decline in efficiency for the silicon regulator. The maximum efficiency was 74.7% compared with a maximum of 76.5% for the unit using germanium chopping transistors. By comparing Figures 8 and 11 more closely, it can be noted that the breadboard with the germanium regulator had an efficiency about one to one and a half percent, higher than the circuit using an "all silicon" regulator. Thus, these experiments have shown that the regulator using germanium transistors has a slight but insignificant efficiency

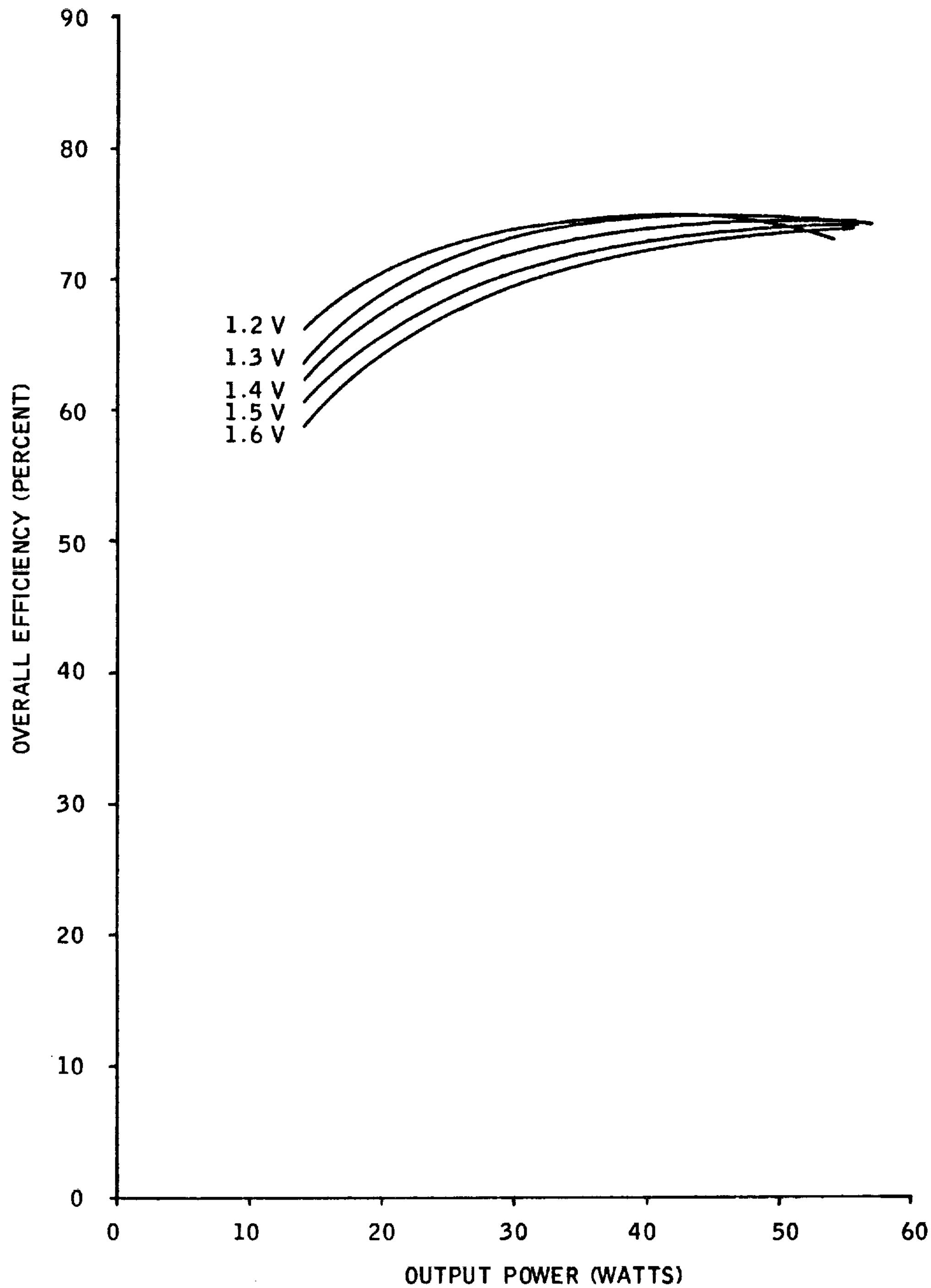


Figure 11 - OVERALL-EFFICIENCY VS OUTPUT POWER FOR A CONVERTER WITH AN ALL SILICON REGULATOR

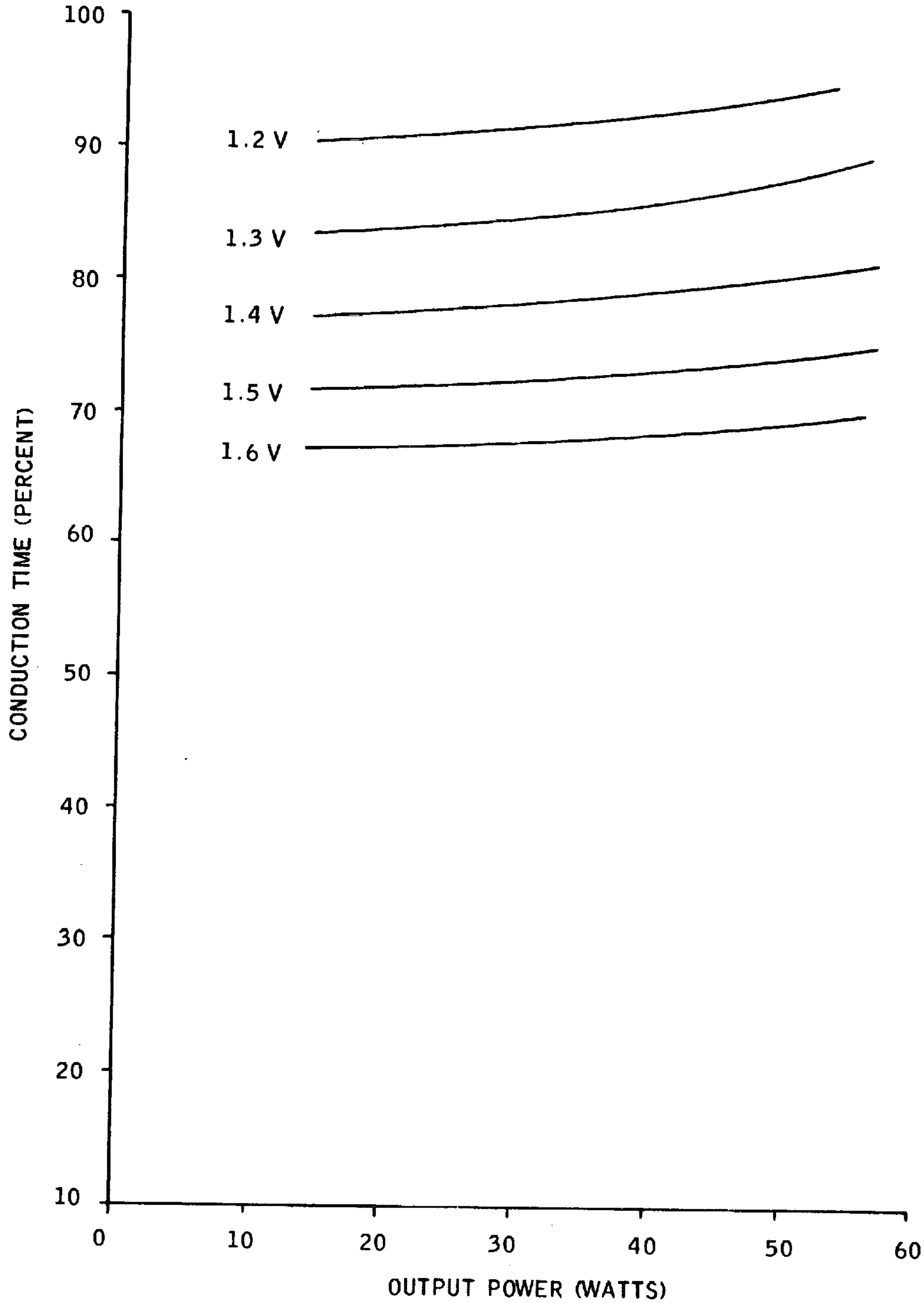


Figure 12 - PERCENT CONDUCTION TIME VS OUTPUT POWER FOR AN ALL SILICON REGULATOR

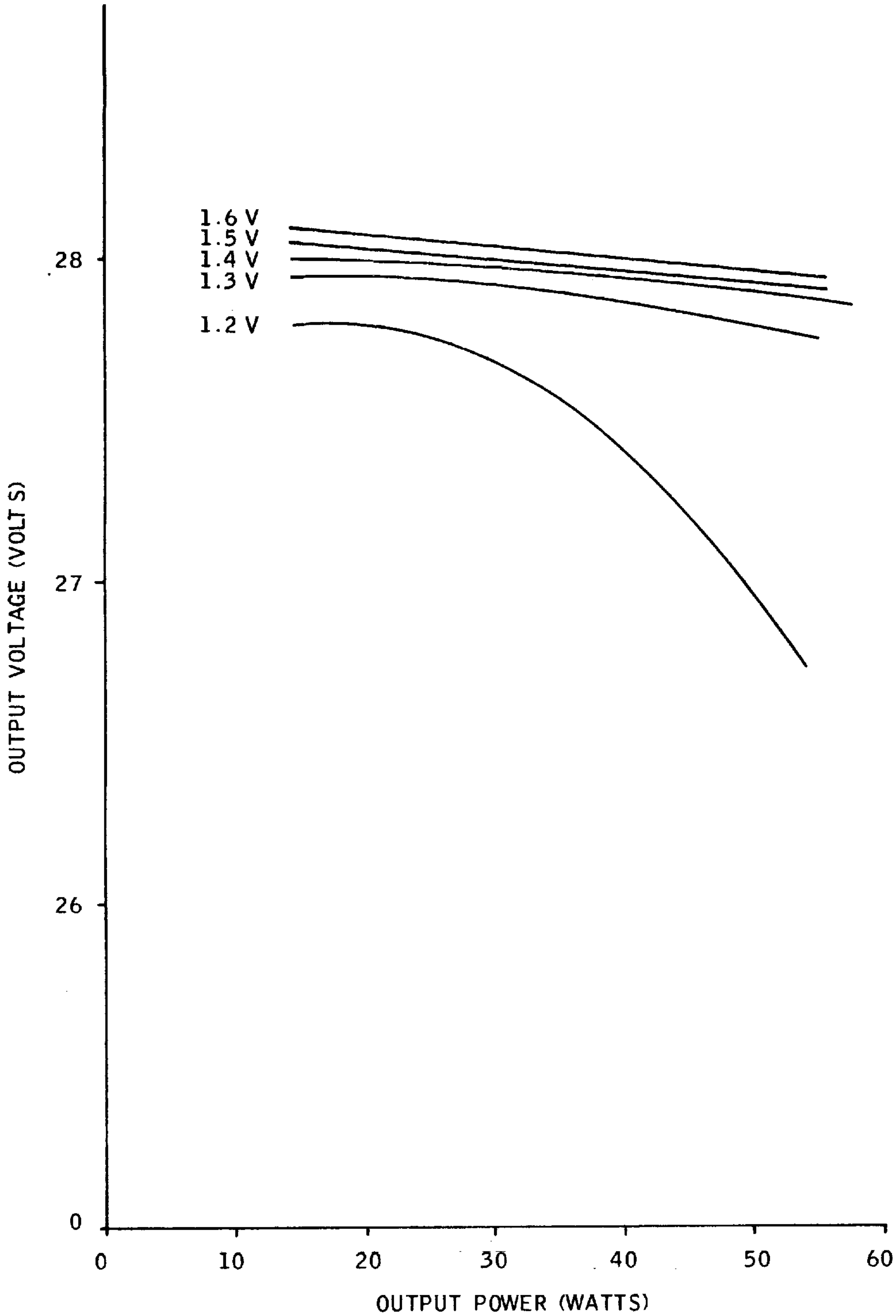


Figure 13 - VOLTAGE REGULATION VS LOAD FOR A CONVERTER WITH AN ALL SILICON REGULATOR

advantage. Therefore, either 2N2833 germanium or Honeywell MHT 8003 silicon transistors can be used for the pulse width modulation regulator. It should be pointed out that two 2N2833 germanium transistors were used in parallel, whereas the "all silicon" regulator used only one MHT 8003 silicon transistor. If only one 2N 2833 transistor were used, the "all silicon" regulator may have shown superior performance. The silicon regulator transistor "Percent Conduction Time vs Output Power" is shown in Figure 12. This family of curves is uniformly spaced and is very similar to the curves for the germanium regulator shown on Figure 9. No significant differences can be noted.

Voltage regulation of the breadboard using the all silicon regulator is shown on the performance curves on Figure 13. For 1.3 to 1.6 volt inputs, the output voltage remained reasonably flat and linear with load. It can be noted that the straight line curves of Figure 13 have slightly more droop with load than the corresponding regulation curves for the germanium regulator shown on Figure 10. Thus, the silicon regulator has slightly more droop. The silicon regulator curves for the 1.3 to 1.6 volt input show that the output voltage remains within the ± 1% required. The 1.2 volt curve however shows considerably more droop at heavy load. This droop occurs when the regulator transistor "Per cent Conduction Time" approaches the maximum limit (over 90%), and output voltage droops due to lower gain and system IR drops.

5. Measured Regulator Efficiency

The performance of the "all silicon" regulator was checked by placing an ammeter and a voltmeter in the secondary circuit between the converter output and the regulator input. This instrumentation measured the regulator input power, except for the relatively small regulator drive taken from transformer T_3 winding N6 shown on Figure 2. The regulator efficiency was then obtained by dividing the output power by the measured input to the regulator. The results were plotted on Figure 14, which shows "Measured

Silicon Regulator Efficiency vs Power Output" for various input voltages. The family of curves shows that higher regulator efficiencies are obtained with lower input voltages. With a 1. 2 volt input the measured efficiency ranged between 93. 8 and 97%. The efficiency is higher at this voltage because the regulator conduction time is very near the maximum. With a 1. 3 volt input the measured efficiency ranged between 89% and 95%. At the 1. 6 volt input the measured efficiency ranged between 85. 9% and 93. 3%. The efficiency declines at higher input voltage and light load because the percent of regulator conduction time declines. Also, at light load the fixed regulator losses are a greater percentage of the output.

The MHT silicon transistor was found to work quite well in the above regulator. An oscillograph of this transistor operating in the regulator circuit is shown in Figure 15. This trace shows a plot of V_{CE} vs time. The converter was operating with a 1. 4 volt input and a 55 watt output. The vertical scale represents 10 volts per division, and the horizontal scale represents 100 microseconds per division. Note that the switching time is very fast. The lower portion of this oscillograph trace was expanded to show the saturation voltage of this transistor when operating in the regulator. The expanded trace is shown on Figure 16. The vertical scale represents 0. 5 volts per division. Note that the MHT 8003 silicon transistor saturation voltage was approximately 0. 1 volt when the regulator output was 55 watts. (The output current was two amperes). The above performance data indicates that this high frequency, low saturation voltage silicon transistor compares very favorably with high frequency germanium transistors in switching regulator applications. The silicon transistor will have an advantage of higher temperature capability. Further work will be done on the voltage regulator circuit to determine if germanium or silicon transistors should be used for the switching element.

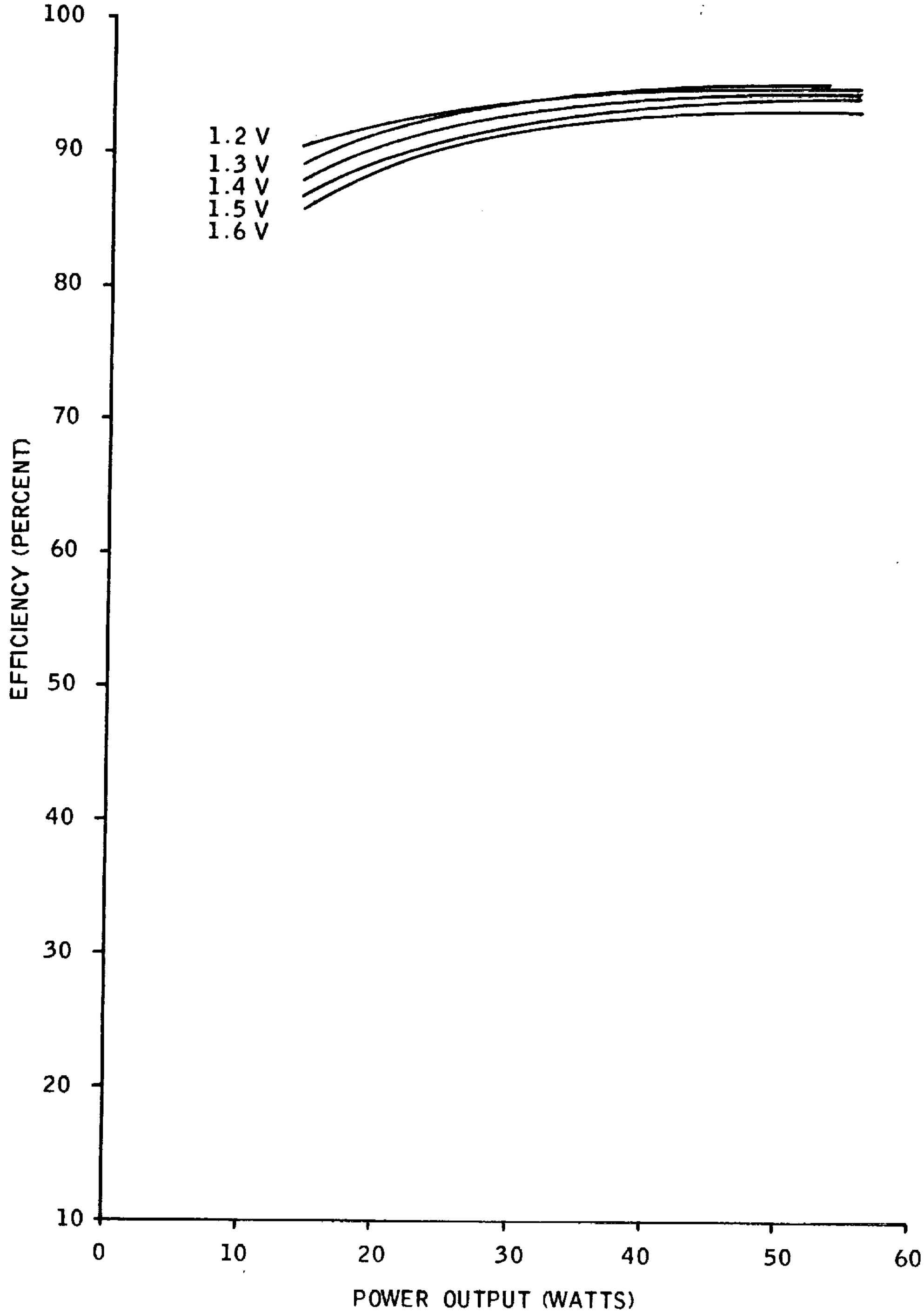


Figure 14 - MEASURED SILICON REGULATOR EFFICIENCY VS POWER OUTPUT

Figure 15 - V_{CE} WAVEFORMS OF THE MHT8003 TRANSISTOR WHILE OPERATING IN THE REGULATOR CIRCUIT

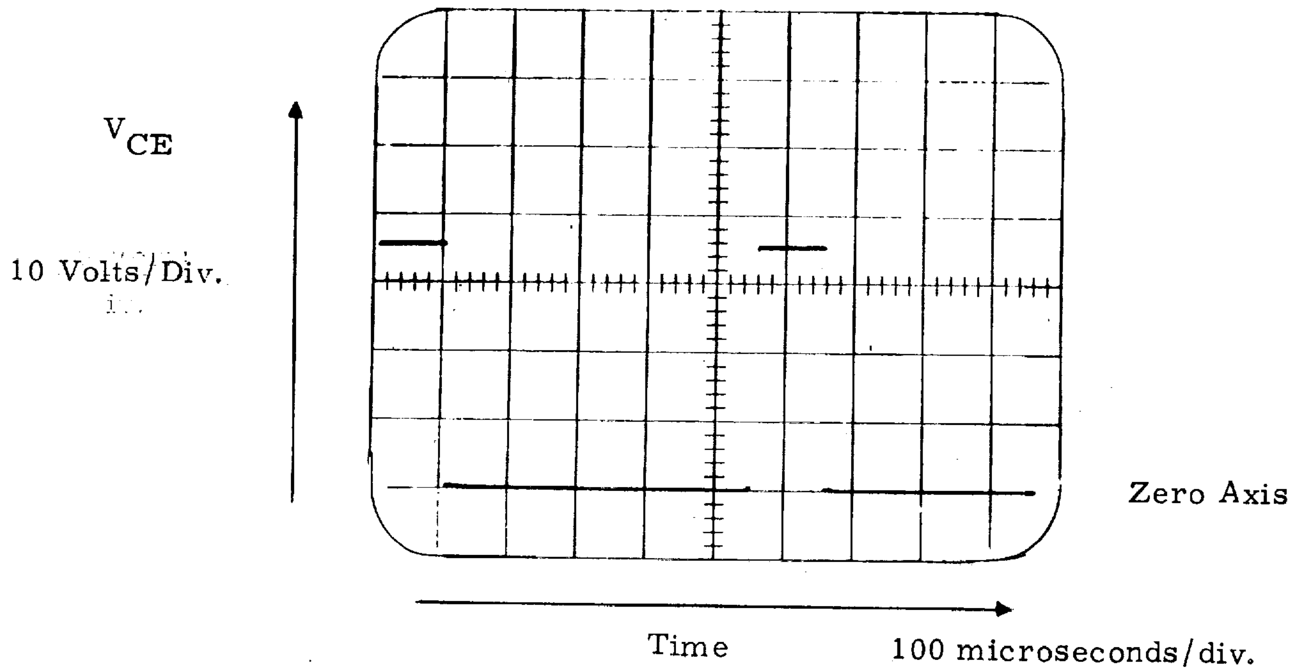
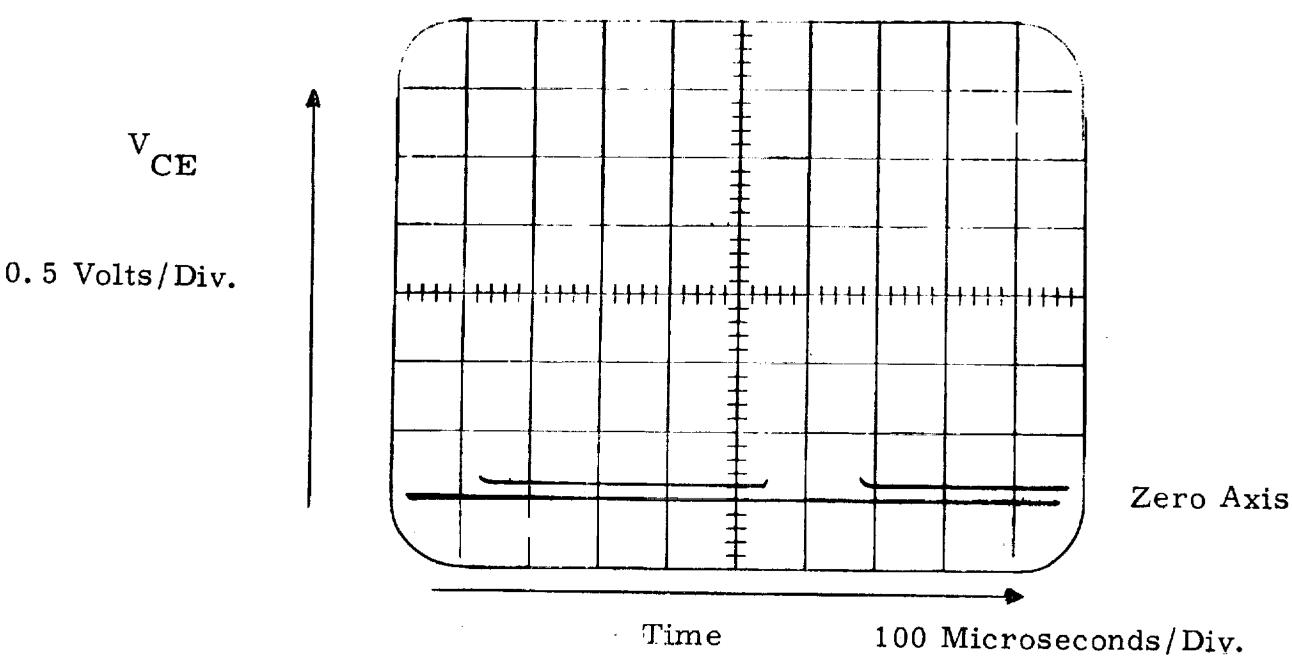


Figure 16 - EXPANDED SCALE SHOWING SATURATION VOLTAGE OF SWITCHING REGULATOR TRANSISTOR



6. Converter Operating Frequency

The converter operating frequency increases as the input voltage is increased. The circuit arrangement to accomplish this is described above under the "Circuit Diagram" heading. Figure 17 shows the converter operating frequency plotted against output power for various input voltages. The curves show that the operating frequency increases with higher input voltages and declines slightly with load. These characteristics are desirable because a higher operating frequency will reduce the operating flux density, and this tends to hold the flux density constant as the input voltage is increased.

Limiting the flux density to a safe value will prevent saturation of the output transformer. Maintaining the operating flux density considerably below B (max) is desirable because a small amount of half cycle unbalance can cause saturation of the output transformer toward the end of one half cycle. This saturation would cause high transistor switching losses. Thus it is desirable to sense the operating flux (or transformer voltage time integral) to control the operating frequency. This sensing is done by reactor L_1 in Figure 2. With the proper choice of transformer parameters the present converter-regulator circuit has been found to operate at a satisfactory frequency without evidence of half cycle unbalance.

The present circuit operates between 515 and 775 cps. Further work will be done to increase the operating frequency. The use of other transformer steels such as "48 alloy", "Hy-mu-80," and "Supermalloy" will be considered. More optimum arrangements for the transformer primary leads may reduce primary leakage inductance, which should improve the converter switching characteristics.

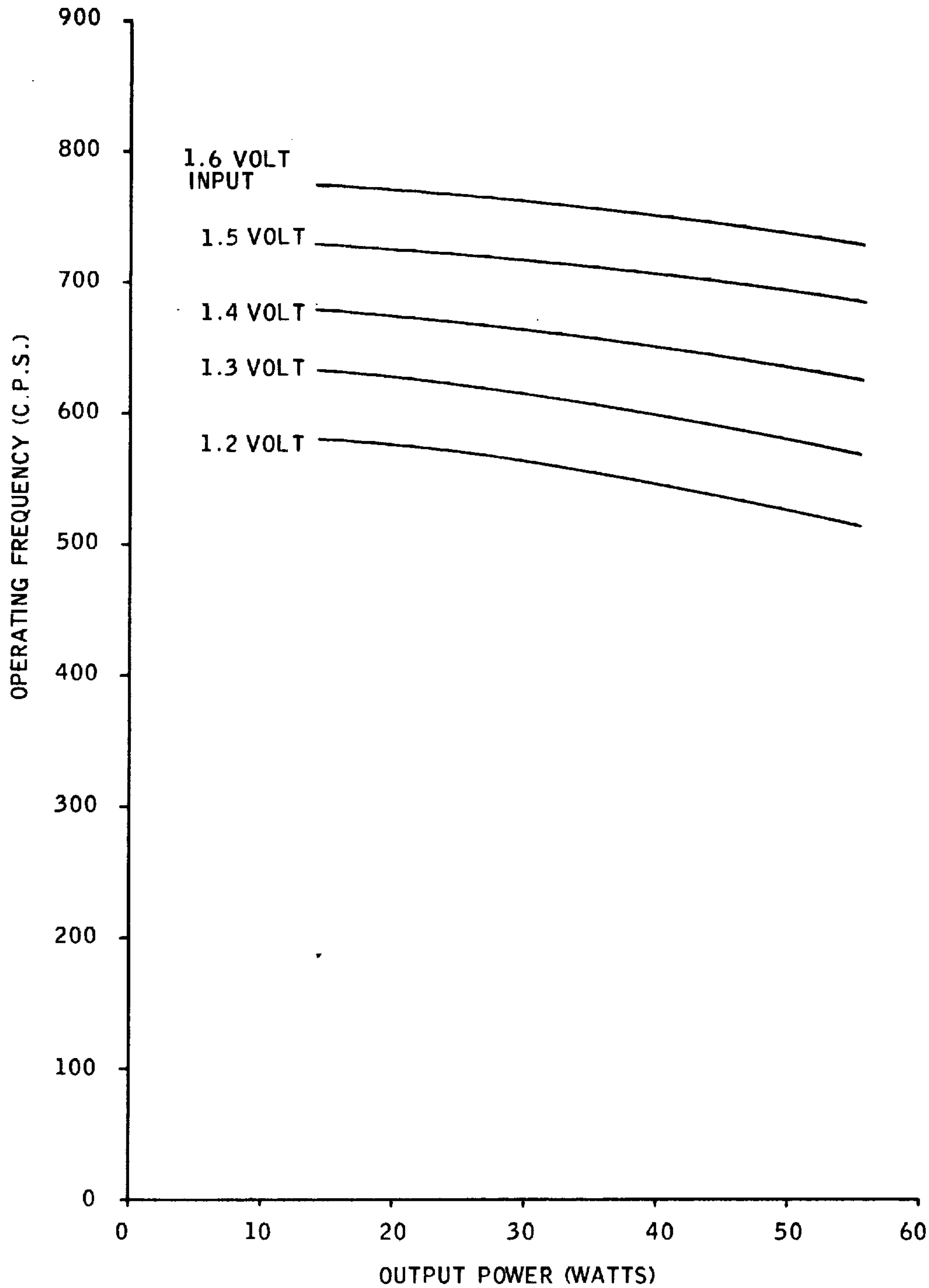


Figure 17 - CONVERTER OPERATING FREQUENCY VS LOAD FOR VARIOUS INPUT VOLTAGES

SECTION V

CONCLUSIONS

The continued effort during this quarter has simplified the converter circuitry and has resulted in improved performance. The simplified converter section reduces the number of parts, improves reliability, and minimizes size and weight. The current feedback power oscillator has been modified so that the operating frequency is proportional to input voltage. This proportional frequency control maintains the power transformer maximum flux density nearly constant and prevents saturation of the output transformer. Maintenance of the operating flux density within specific limits prevents saturation of the output transformer due to minor half cycle unbalance, assuring operation at maximum efficiency.

Preliminary ambient temperature tests have verified that operation is satisfactory over the +70°C to -10°C temperature range. The converter-regulator was checked at temperatures much lower than required (-54°C) and was found to function satisfactorily. The regulated converter efficiency increases slightly at lower temperatures. The temperature tests have established that the regulator can be compensated for temperature variations so that the output voltage will remain within ± 1% over the +70°C to -10°C operating range. Because the converter has been modified since the preliminary temperature tests, it will be necessary to perform additional temperature tests to verify that the regulator temperature compensation and input voltage compensation are satisfactory.

Performance data (Figure 8) shows that during the past three month period, the regulated converter efficiency has been increased from the 72% level to the 74.5 to 76.5% level with a 50 watt output. Measurements indicate that further increases in efficiency may be achieved by eliminating the flexible leads used to connect the transistor emitters to the power transformer primary winding. Selection of transistors with slightly lower saturation voltages may also improve performance.

The voltage regulator holds the output voltage within the ± 1% required over the load range with input voltages between 1.2 and 1.6 volts. Examination indicates that additional input voltage compensation would improve the regulation further. The regulator output voltage will begin to droop as the regulator percent of conduction time approaches the maximum limit (over 95%). The regulator efficiency will decline as the percent of regulator conduction time diminishes due to higher input voltages or reduced loads. Therefore, it is desirable to select an optimum input voltage range for a particular source-converter-regulator arrangement so that the efficiency can be optimized over the anticipated load range. It would be desirable to study the performance of a boost-type regulator that pulse width modulates only a portion of the power and compare its performance with the present circuit which pulse width modulates all of the power. The efficiencies, sizes, weights, and complexities could be compared to determine what the trade-off possibilities are.

Performance data for the voltage regulator using one of the new Honeywell MHT 8003 high frequency low saturation voltage silicon transistors has shown performance which compares very favorably with the regulator using two parallel connected germanium transistors as the switching elements. The difference in overall efficiency between the two is only one or two percent. These results indicate that an "all silicon" regulator can be considered for this application. This regulator would relieve heat transfer problems in the regulator section.

Experiments have shown that the overload protection circuit protects the converter from both slowly rising overloads and from short circuits. During overload the output current is cut back to a low value which is sufficient to monitor the output and determine if the fault continues to exist. When the fault is removed, the overload protection circuit automatically deactivates and normal output power will be supplied to the load.

Work during the past three months has refined the circuitry to a high degree. Effort must now be concentrated on model design to minimize the size and weight of the device. If the present breadboard circuit components were packaged, it is accurately estimated that the weight would be six pounds and the size 150 cubic inches. Further design efforts can reduce the component size, and the four pound weight and 120 cubic inch size required still seem feasible. Weight reduction will require more effort than size reduction. Honeywell will work on these problems during the next three month period and will fabricate a breadboard model which will be delivered to NASA.

SECTION VI

PROGRAM FOR THE NEXT INTERVAL

During the next interval the main effort will be concentrated on the design and fabrication of a breadboard model to be delivered to NASA. This work will involve the design of the main components, transformers and chokes, as well as the design of the breadboard package. Further temperature tests and performance tests will be conducted during this interval to verify the selection of all circuit parameters required to achieve proper temperature and input voltage compensation of the regulator. Additional tests will be made to determine if the regulator chopping transistor should be germanium or silicon.

Emphasis will be placed upon size and weight reduction. The use of electroplated aluminum bus bars for the input connections and transformer primary will be considered to reduce weight. The portion of the primary winding that passes through the center of the torroidal power transformer will probably be copper, however. Special effort will be made to minimize the lead resistance between the transformer primary winding and the oscillator transistors. Solid, soldered connections will be used in place of flexible leads. A conference will be arranged to discuss the program and model fabrication with NASA personnel.

The breadboard model will be given environmental tests and a preshipment test before being shipped to NASA.

Prepared by John T. Lingle

John T. Lingle

Approved by B C Tierney

B. C. Tierney

D. A. Nelson

D. A. Nelson

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